

INTERMODULATION

DYNAMIC RANGE

AND PHASE NOISE

THE CONCEPT of intermodulation as a limiting parameter in receiver performance has become more common over the last few years, possibly because of the introduction of wideband solidstate receivers in which the performance in this respect was worse than that of the valved receivers they replaced. More recently, the term "dynamic range" has been introduced as a catch phrase to encompass the inadequacies (or otherwise) of receivers and radio systems in what is frequently a non-quantifiable manner. As a result of this fashion, the term "dynamic range" now means all things to all men, and almost any meaning desired can be attributed to it!

However, by specifying the system in which we are working, some of the confusion can be removed. For example, the "dynamic range" of an fm i.f. strip is generally considered to be the range of on-tune signals which can be handled without excessive distortion or damage. In a radar receiver with a logarithmic i.f. amplifier, the concept of dynamic range as that range of signals over which the logarithmic law holds is valid. But in a communications receiver the situation is clouded by the very environment in which the receiver must work, ie in the presence of a large number of unwanted signals. At first glance, it may well appear that the problem of handling strong signals is easy, for an effective agc loop will reduce the gain of the receiver as desired to remove the overload. This is, of course, perfectly valid—for signals within the detector passband of the receiver. But for signals that are off-tune, a completely different situation occurs, because a signal which is outside the detector passband of the receiver will not produce any agc, and so the overload still occurs. Therefore the agc dynamic range is a different parameter to the dynamic range, which is that of the ratio of unwanted signal on an adjacent channel to wanted signal for a given degradation in received signal quality—while this is unlikely to bear much resemblance to the "spurious free dynamic range" used in discussing intermodulation. It is therefore possible to consider the term dynamic range as also covering a composite measurement of receiver performance, although this has not yet been produced.

In the conventional, non-synthesized receiver, especially if solidstate, the major limitation in performance is produced by intermodulation, and it is therefore reasonable that this should be considered first. (For a discussion of receiver parameters and their meanings, readers are referred to [1]).

As a phenomenon observable by the amateur, intermodulation has been observed in transmitters since ssb was first introduced. In its simplest form as it occurs in an ssb transmitter under two-tone modulation, it is fairly easy to understand, and reference to Fig 1 will make the picture clear. Here we have an amplifier in which the input consists of two frequencies, which for the sake of convenience are between a few hundreds of hertz and a few tens of kilohertz in separation. The effect of non-linearity in the amplifier is to produce at the output of the stage a series of harmonics of the input frequencies—so, for example, if the input frequencies were 7,000 and 7,001kHz, the output would have components at 14,000, 14,002, 21,000, 21,003, 28,000, 28,004 and so on. Were these all the output components, the problem would be simple, for the unwanted components would be removed by simple filtering. Unfortunately, however, this is not the case, because devices which will produce harmonics will also act as mixers.

The result of this is that the various harmonics mix with each other to produce a number of spurious outputs, which are known as intermodulation products. For example, the second harmonic of 7,001kHz is 14,002kHz, and this mixes with the fundamental of 7,000kHz to produce an output component of 7,003kHz. Similarly, the second harmonic of 7,000kHz is 14,000kHz, and this mixes with the 7,001kHz to produce 6,999kHz. Note that these spurious outputs are very close to the wanted output frequency, and therefore cannot be removed by filtering. In a similar manner, combinations of $3 \times 7,001$ minus $2 \times 7,000$ produce an output on 7,003kHz. Generalizing, then, if we have two signals on frequencies f_1 and f_2 , the output intermodulation products will be on frequencies given by the equations $mf_1 - nf_2$ and $mf_2 - nf_1$, where m and n are integers; for example, products of $2f_1 - f_2$ and $2f_2 - f_1$ are typical examples. Where the sum of m and n is even, the resulting products are considerably removed in frequency from the fundamental input frequency, and are thus fairly easily filtered. These products are referred to as even-order products, and may generally be ignored in practice.

However, where the sum of m and n is odd, inspection shows that a series of outputs is produced which is closely grouped around the wanted output frequency. The spacing between the signals (often referred to as imps, intermodulation products, or just as ips) is the same as the spacing of the two fundamental signals, and the lower the sum of the values of m and n , or order of the ip, the closer those products are to the fundamental. It is ips of this sort which are so objectionable to other spectrum users, and lead to complaints of splatter or broad transmissions. Interestingly, the level at which ips begin to become audible in terms of distortion of the received signal is considerably higher than that at which severe annoyance is caused to the occupants of the adjacent channel. While with speech transmission, statistical effects allow a much worse ip level to be used than would at first be expected from a study of adjacent channel power.

It is at this stage that the reader may well be wondering about the significance of this parameter, normally associated with transmitters, in the context of receivers. Also, the purist may well be complaining that the above explanation of intermodulation is incorrect—after all, if the ips are caused by the mixing of harmonics, why can they not be trapped out? The answer to the latter question is, of course, that a simplified explanation of the mechanism has been given, which fits the observed phenomena reasonably well; but without a rigorous mathematical analysis, the distinction in mechanisms is neither easily explicable nor particularly useful. Nevertheless, it should be noted that the mixing of harmonic components actually appears within the device, and the individual currents involved at the various frequencies are not available to allow filtering to take place.

If we now consider the simple case of a receiver tuned to a weak signal on 7,010kHz, with strong signals being present at the input of the receiver on

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In September 1983 he married G4FNC, daughter of G3ZCV and sister of G4M2M, thus keeping amateur radio firmly in the family! His other interests include railways, canals, cooking and home-brewing (beer as well as equipment).



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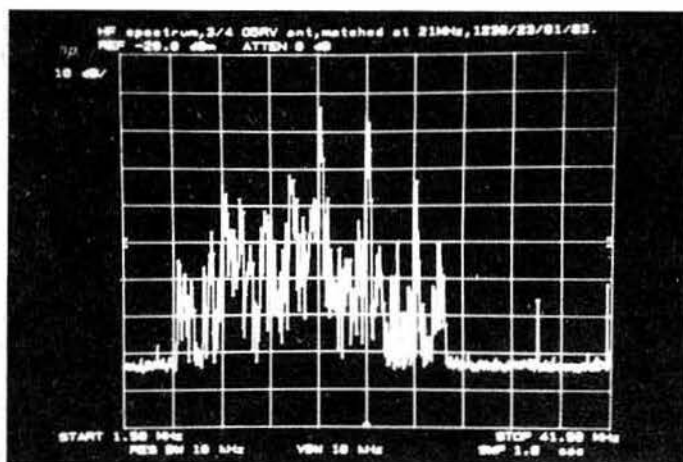


Fig 1. The hf spectrum, 1.5 to 40MHz

7,020 and 7,030kHz, the effects of third-order intermodulation will produce an ip on 7,010kHz—which is the frequency of the weak signal which the receiver is trying to receive. It is at this point that the applicability of intermodulation to the receiver is apparent. However, the simplistic approach of considering two signals, while conveying very well the basis of the problem, does not begin to address the real world in which the hapless hf receiver has to operate.

White light is made up of all the frequencies in the optical spectrum, and the radio frequency spectrum—if looked at over a wide enough bandwidth—similarly appears to be full of noise. If we were to include random amplitude and phase enough discrete frequencies, we would produce a signal indistinguishable from white noise, and even the addition of the signals in part of the hf band produces an approximation to noise. The receiver, by virtue of its selectivity, is able to sort out the desired signal from the noise, in the same way that red glass in the traffic light filters the red light out from the white light generated in the bulb.

However, if the linearity of the front-end of the receiver is such that intermodulation can occur, then the vast numbers of intermodulation products which can be produced from a number of signals produces an output closely akin to noise. The more mathematically inclined may wish to list the output ips from three input signals up to the seventh order as an exercise in the possible problems. So the front end of a completely wideband receiver is quite likely to be heavily overloaded by the output from even a modest antenna, while the connection of, say, a rhombic may well lead to few signals being receivable. The use of an antenna attenuator to cure this problem has been suggested, and although this can be of assistance, the author considers this approach to be an admission of defeat; insofar as in a properly-designed receiver, such an attenuator is unnecessary unless high-power transmitters in the immediate vicinity of the receiver are a problem. Even so, it is necessary to achieve some understanding of the levels of intermodulation products and their relationship to the power being handled in order to appreciate this relationship. For cw, the requirements are much more severe; eg -94 dBc/Hz at 600Hz.

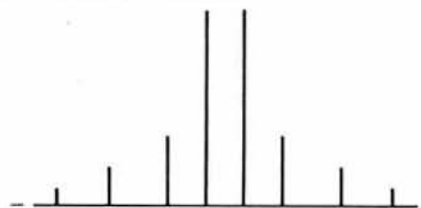


Fig 2. Two tones and intermodulation products

In a perfectly linear system, the relationship between input and output is linear; if the input power is doubled then the output power is also doubled. Obviously, some limit must exist, or we would have BC108s producing 500W, which is patently absurd. Fig 3 shows this relationship in graphical form, and it can be seen that, in practice, as the limit of output occurs, the gain of the system decreases. Where the gain has decreased by 1dB, the point reached is the "1dB gain compression point", and the system generally has reached gain compression. The actual and theoretical curves in Fig 3 show this change in performance. Note particularly that the gain compression point can be specified in terms of either input or output power at the compression point, and which is used is often a matter of "specmanship".

The level of intermodulation products in a system operating below the gain

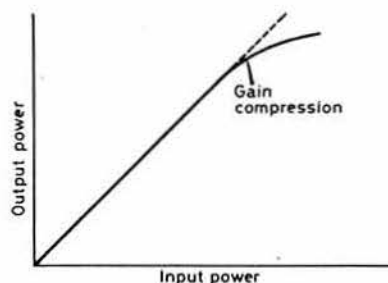


Fig 3. Gain compression and input/output characteristics

compression point is dependent upon the input power to the system, and the order of the intermodulation product. For every 1dB change in input power, the third-order ip will change by 3dB, the fifth-order by 5dB, and so on. So we can establish a family of curves as in Fig 4, where the various lines show the rate of rise in the ips with signal. Usually, for the sake of clarity, only the third-order products are shown, for these are the most troublesome.

So we can see that in the case of the 7MHz receiver discussed earlier, the addition of 1dB of attenuation in the signal path will attenuate the wanted signal by 1dB, but will attenuate the ips by 3dB, which is a net gain of 2dB when the noise floor of the receiver has been artificially raised by intermodulation effects. It is for this reason that antenna attenuators have become popular, but it is of course quite obvious that in the ideal case the attenuator would not be necessary because of the adequate intermodulation distortion capabilities of the receiver. So it is a fair argument to suggest that the inclusion of an attenuator is either an admission of defeat in design capability, or a pandering to a somewhat ill-informed whim.

A popular term used in intermodulation discussion is that of intercept point. Fig 4 shows this better than any description can; but, briefly, intercept point is the purely theoretical result obtained by extending both the input against output line and the intermodulation against output line to the point where they intercept or cross. This point is necessarily theoretical because the component under discussion would be well into gain compression long before the power of the ip was equal to the power of the fundamental: nevertheless it provides a very convenient method of determining performance. As in the case of compression point, there are two intermodulation intercept points, the input and the output intercept points, and they differ by the gain below the compression point. The usual intercept point quoted for components such as amplifiers and mixers is third intercept point, although in wideband systems, the second-order intercept point is very important. Nevertheless the adoption of a push-pull configuration is very effective in minimizing second-order intermodulation problems.

It is the second-order products which lead to the inclusion of "sub-octave" filters in hf receivers. Second-order intermodulation is of the form $f_1 + f_2$ or $f_1 - f_2$ producing the wanted signal: eg $11\text{ MHz} - 4\text{ MHz} = 7\text{ MHz}$. It can be shown by a simple arithmetical exercise that if the bandwidth of a system is less than one octave, second-order intermodulation cannot occur. The approach of using a high i.f. and upconverting, and using a lowpass filter to provide image rejection has proved particularly prone to suffer in this respect, as has the use of very wideband antenna amplifiers.

Obviously, intermodulation products will only be a problem if they are above the noise floor of the system, and it is comparatively easy to show that signals which exceed the noise floor by less than two-thirds of the amount that the intercept point exceeds the noise floor will not cause trouble. It is this range of signals that is usually understood as being concerned in the meaning of dynamic range. Here an example may usefully be used.

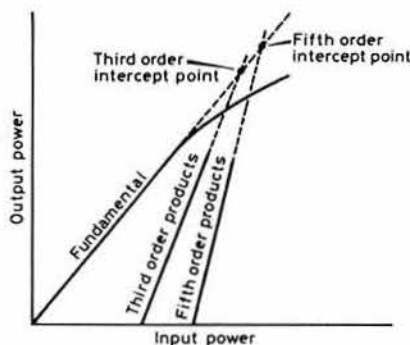


Fig 4. Intercept point

A receiver has a 20dB signal-to-noise ratio for a 1μV input. Two signals each at 1mV produce a 1μV intermodulation product. It is usual to express intercept points in "dBm" (decibels relative to 1mW in 50Ω). This nomenclature has some advantages insofar as for those not accustomed to working in decibels, the conversions can prove somewhat easier: in addition, certain system calculations are eased. 1μV in 50Ω is -107dBm, so the receiver in our example has a noise floor 20dB below the signal, or at -127dBm. We also know that two signals at 1mV (-47dBm) produce a 60dB intermodulation ratio. Extending this figure, bearing in mind the rules outlined above, leads us to say that at -37dBm the intermodulation ratio (imr) will be 40dB; at -27dBm input it will be 20dB, and at -17dBm input the imr will be 0dB. However, the point at which the imr is 0dB is the intercept point, and we can thus say that this receiver has a third-order input intercept point of -17dBm and a noise floor of -127dBm. With this intercept point, any signals lower than -60dBm (220μV) will not produce signals above the noise floor of the receiver—see Fig 5. From this, we therefore have a dynamic range of 2/3 (-127 - (-17)), or, neglecting the minus sign, 73-66dB. In other words, where signals can exist at the receiver input at a level greater than that given by noise floor (dBm) + dynamic range, intermodulation troubles can be expected to occur. This dynamic range, which recently started appearing in the specifications of amateur receivers, is correctly known as the "spurious free dynamic range", and is complicated by the fact that it is nowhere near as easy to measure as might at first be thought [2]. Note that the use of an antenna attenuator does not increase the dynamic range, but only moves the effective noise floor of the receiver.

The topic of intermodulation has been extensively covered in the literature, and [3] and [4] may well be of interest to the reader.

In summary: intermodulation can limit the effectiveness of a receiver by providing not only discrete spurious signals, but by raising the noise floor because of the products of the vast number of signals present at the input. Dynamic range in this sense is the amount by which signals must exceed the noise floor for intermodulation to be a problem.

Phase noise, reciprocal mixing and dynamic range

If dynamic range is alternatively defined as the ability of a receiver to respond to a small wanted signal in the presence of a number of large unwanted signals, other effect becomes apparent. The effect is known as reciprocal mixing, and is caused by phase noise in the local oscillator of the receiver. It has only become of much importance in the last 10 or so years, merely because the techniques used previously performed very well, and again it was the advent of the solidstate synthesized receiver with its drop in performance that has led to the importance of reciprocal mixing being recognized in professional circles.

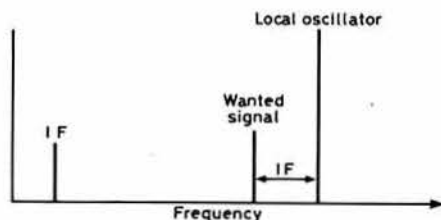


Fig 5. Clear spectral lines in mixing

Fig 5 shows the relationship in the usual superhet between the input, oscillator and intermediate frequencies, and shows how the effective "window" of the i.f. passband is tuned across the input frequency range by varying the local oscillator frequency. A large signal, offset in frequency by some amount, is rejected by the i.f. filters, and causes no problems with the weak wanted signal. In Fig 6 we see the effect of deliberately frequency-modulating the oscillator with noise, so that instead of being a spectral line, it becomes a broad "block", and is thus able to mix any signal within the block to the i.f. Obviously this is a case of *reductio ad absurdum*, for no-one would deliberately modulate the oscillator in such a way. However, the practical case is such that the oscillator is modulated with noise at some level, and in practice the situation is much more akin to that shown in Fig 7, where there is some noise modulation of the local oscillator. It is this broadening of the signal at low levels which gives rise to the problem, while the effect produced, that of noise mixing with the unwanted signal, gives rise to the name reciprocal mixing. In terms of the selectivity of the receiver, it means that the ultimate rejection of the i.f. filters can well be impaired,

while the performance required to give reasonable overall receiver performance is quite high and difficult to achieve. The noise modulation produced is not a.m. noise, which, theoretically at least, is fairly readily removed, but is phase noise, or phase modulation produced by noise. There are various techniques for reducing phase noise, some of which will be mentioned later. However, some degree of quantitative measurement is required, and this will now be discussed.

The textbook introduction to noise shows us that the power produced in a resistor by the random movement of electrons is given by

$$P = kTB$$

where P is in watts, k is Boltzmann's Constant (1.37×10^{-23} J/K), T is the absolute temperature, and B is measurement bandwidth in hertz.

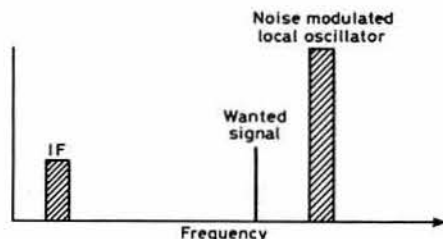


Fig 6. Noise modulation of the lo

By "plugging-in" the figures for a 1Hz bandwidth at room temperature, we obtain a figure of -174dBm as the absolute minimum limit that a noise floor can be at in a 1Hz bandwidth. In a 10Hz bandwidth this will be -164dBm, and so on *pro rata*. A perfect oscillator running at an output of 1mW (0dBm) can therefore never have a better signal-to-noise ratio than 174dB, and so the advantages of an oscillator running at high power become obvious. In an AR88, for example, the input to the oscillator is about 1W, and the signal-to-noise ratio with a 10 per cent efficiency is then limited to some 194dB maximum—a figure which is reduced in the practical case by the noise figure of the valve and the finite Q of the tuned circuits and other components. Nevertheless it is quite obvious that the restriction of input power to an oscillator in the interests of stability is at variance with the best phase noise performance.

The noise in an ssb bandwidth of 3kHz is some 34dB above noise in a 1Hz bandwidth, which is a convenient figure to use for comparison purposes. If the response of the i.f. filters is specified down to 60dB, then obviously the phase noise of the oscillator must be better than 60dB in the particular bandwidth if the filter is not to be effectively degraded. Translating this back to the reference 1Hz bandwidth, the oscillator phase noise limit is 94dBc, which means -94dB relative to the carrier power of the oscillator, when measured in a 1Hz bandwidth.

It is at this point that the question may well be raised "If the noise is phase noise, why is it detected in a linear detector such as the product detector in an ssb signal?" The answer to this is that the ssb signal contains phase and amplitude information; variation of the frequency of an ssb signal obviously affects the output of the receiver, and the phase noise introduced by the local oscillator chain may well be considered as fm.

To recap: the noise at a given spacing from the local oscillator (lo) signal will be at some level X decibels down on the lo signal in a given bandwidth. An unwanted signal X decibels above the wanted on-tune signal will mix the noise of the local oscillator into the i.f. bandwidth, the unwanted signal being distant in frequency by the offset from the lo frequency.

So in a receiver where some £30 has been spent on a KVG XF9B filter with a stopband of 90dB, the lo noise has got to be -90dBc in a 2-1kHz bandwidth (-123dBc in 1Hz) if the stopband is actually to be realized in practice. For this reason the claims of 120dB ultimate rejection used in the advertising for the Atlas transceivers some years ago was meaningless when viewed in terms of receiver performance. References [5] and [6] show the results of reciprocal mixing on two modern hf transceivers in a very effective manner.

It has now been shown that the capability of a receiver to do its job is likely to be limited, especially at hf, by either intermodulation or reciprocal mixing, or both. Furthermore there would not appear to be any particular virtue in producing a receiver with outstanding performance in one respect without taking necessary precautions to obtain good performance in the other.

It now remains to decide what levels of performance are required in a receiver, and under what conditions. For this exercise the receiver considered will be for the hf bands only, but exactly the same considerations can be made for a vhf receiver—and at this point the writer is prepared to stir up some controversy by saying that the vhf receiver is no more demanding in dynamic range terms in general than is the hf one.

References [6] and [7] give a very complete picture of the hf band for general coverage receivers, but some modification of these figures may well be desirable for amateur use, insofar as the antennas used are either inherently narrowband (such as quads, Yagis or dipoles) or else are used with narrowband matching networks (atus) which provide a useful limitation of the input power. Unfortunately the losses in mismatching from an off-tune dipole to the average receiver are not as high as could be hoped, so the adoption of levels of the above references are probably only excessive by about 10dB.

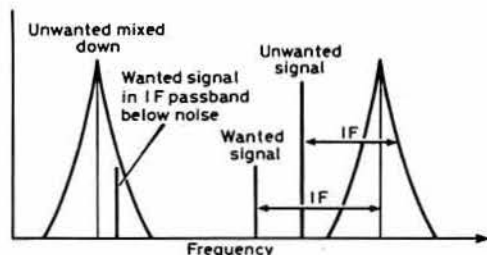


Fig 7. The practical reciprocal mixing case

Barrs [7] deals with the case in which the front-end of the receiver is restricted to a bandwidth of ± 2.5 per cent, representing some 700kHz at 14MHz. Using the case of a receiver with a +24dBm input intercept point, it is shown that there will be only one or two signals as high as 100mV emf within the ± 2.5 per cent band. If we assume two 100mV emf signals, there will be the following intermodulation products:

- 6 up to third order at 30 μ V emf
- 14 up to seventh order at 3 μ V emf
- 30 up to fifteenth order at 0.3 μ V emf

This is using a traffic density pattern which gives the following band occupancy:

Input level (mV)	Frequency (MHz)	12	14	16
100				
30		4	3	
10		5	5	
3		10	9	
1		20	18	
0.3		40	36	
0.1		60	60	

In this table the vertical scale represents the signal strength in millivolts emf, while the horizontal axis is scaled in frequency. The numbers in the columns represent the number of stations in the input range.

It is probably a fair assumption that the average amateur station using, say, a three-element beam at 40ft will receive fewer strong signals throughout this band. If we assume that the signals are reduced by 10dB because of the narrowband antenna arrays used by amateurs, reference to the figures published by Hart [5, 6] show that, with typical intercept points of amateur receivers being some 10dB worse than those used by Barrs [7], the numbers of intermodulation products will be approximately equal, as will the levels. In this respect, reference to Barrs again will show that the noise floor of the FT-1 will, from Hart's figures [5] be degraded, probably to about the extent that a 1 μ V signal will be inaudible.

A valid question at this stage is that of how do these equipments work as well as they do. The answer, as already indicated, is that the amateur receiver has the advantage of having a relatively narrowband antenna, and of being used by many operators who are incapable of deciding whether or not a failure to receive is caused by a receiver fault, QRM or a noise floor increase. In addition, many amateur operators are able to accept very marginal copy—after all, with callsign, name, QTH and report repeated three times, a fair degree of communications redundancy exists! In general it is probably fair to say that the average amateur receiver suffers quite badly from reception failure due to these causes, while the shortcomings in this respect are neither recognized nor understood.

If we assume that Barrs' results—diminished by 10dB to allow for the "amateur" factor—are used, the addition of 10dB to the intercept point will obviously cause a vast improvement in the amount of problems caused by intermodulation. In fact, the 50 signals caused by intermodulation will

drop to about 20 if an intercept point of +24dBm can be achieved, and with these products spread out through a 2MHz band, the probability of interference is obviously very much reduced.

From Barrs we can also see that the noise level in a 300Hz bandwidth caused by galactic and man-made noise is about 0.25 μ V at 14MHz. Experience suggests that to apply a 10dB correction to this is hardly fair, as even a low antenna can easily produce a 2dB increase of receiver noise output, compared with that due to a dummy load, when it is connected. From this it is probably fair to suggest that a sensitivity of about 20 to 25dB signal-plus-noise-to-noise ratio for a 1 μ V input is the most that can reasonably be used—corresponding to a noise factor of about 9–12dB. Curiously enough, this noise factor seems to have been adopted commercially by receiver manufacturers: it has been stated, however, that noise figures as low as 8dB can be used where short whip antennas are in use.

Reciprocal mixing has not yet been considered. If we modify Barrs' figures by the same amount, and use Hart's figures for the FT1, we can see that the noise floor will be degraded by some 3dB because of reciprocal mixing. However, this degradation should be insufficient to cause problems purely because of the level of man-made and galactic noise, if we assume that the strong signals are separated by 300kHz or more from the tune frequency. If that does not apply, for example with spacings of only 100kHz being used, then the situation is much worse, while at 20kHz off tune, a 2mV signal would cause appreciable interference to an otherwise readable signal of 10dB signal-to-noise ratio. The increasing use of solidstate pa stages with their appreciably higher order intermodulation products, does suggest that for ssb these parameters are not so important. In the case of the FT1, the high-order products are some 20dB worse than those of the FT102 with its valve pa. Under these conditions, a station causing sufficient interference to a receiver because of reciprocal mixing would, in the event of the removal of the reciprocal mixing problem, suffer another layer of interference a few decibels down caused by the transmitter imps. In fact, the FT1 would suffer QRM because of reciprocal mixing, while the FT102 would suffer QRM from an FT1 because of transmitter intermodulation. This perhaps makes the maxim that "if I interfere with you, it's your lousy receiver, but if you interfere with me, it's your lousy transmitter" seem not quite so far fetched.

Any queries at this stage as to why the Yaesu FT1 and FT102 have been chosen for these examples may easily be answered by the fact that these transceivers have been reviewed recently, and have had the reciprocal mixing measured.

So have we decided upon the level of reciprocal mixing performance required for an advanced amateur receiver? If we assume a filter stopband of 90dB on ssb, then a noise floor of at least -125dBc is required on the local oscillator at 10kHz offset, and this will hopefully fall to about -140dBc at 100kHz (all in a 1Hz bandwidth). Another 20dB on top of this would be desirable, but would be used relatively infrequently on ssb because of transmitter deficiencies.

A valid argument may well be that the extreme skirt selectivities used in amateur receivers are not in fact particularly necessary, because the adjacent channel is unlikely to be occupied with a signal clean enough for the skirt selectivity to be used.

As an example of this, consider the case of a transmitter such as the FT102 running without speech processing at 100W output. The imps on a two-tone test in the adjacent channel 3kHz away will be around -40dB, and speech will be somewhat better because of the high-peak-to-average-power ratio leading to the average power, and thus the average imd power being less—probably about 25dB less. Under these conditions the average intermodulation power in the adjacent channel will be about -65dB relative to the transmitted channel, and thus selectivity of this order in the receiver is usable. Consider now the case of a filter offering 80dB of adjacent-channel selectivity, and it is obvious that it cannot be used to full advantage. Where the offending transmitter uses speech processing, the averaging effects no longer apply, which may well explain the apparent broadness of stations using speech processors under some conditions.

In order to obtain data on the spectral density at the input of an amateur receiver, some measurements were made. The antenna system is probably best described as a "three-quarter 5RV", at a height of 25ft at one end, falling to 16ft at the other. An atu of the W1FB "ultimate transmatch" type was used to provide matching, and the system was tuned for minimum swr on 7,050kHz on a winter's night. Twenty-two signals exceeding 250 μ V were noted in a by-no-means-extensive search from 7 to 7.45MHz, of which five were between 3 and 10mV, eight between 1 and 3mV and eight between 300 μ V and 1mV. In a receiver with an intercept point of 0dBm, these would give 20 imps of 50 μ V or over, 56 of 5 μ V or over, and another 56 of 0.5 μ V or over—a total of 132 spurious signals. This takes no account of the higher order (fifth and seventh) products, or the strong signals on the If side

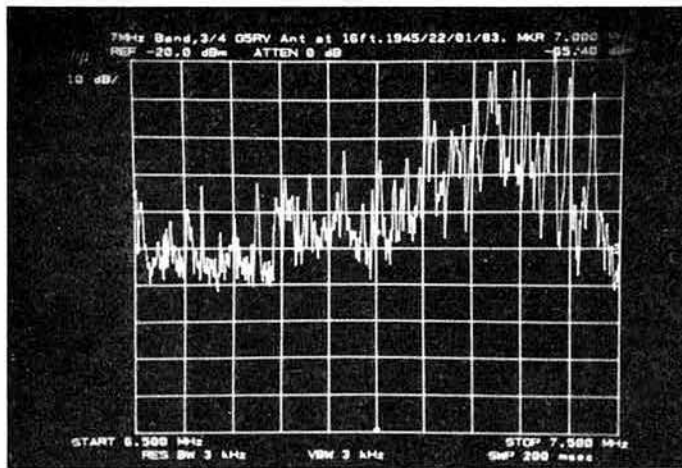


Fig 8. 7MHz at night

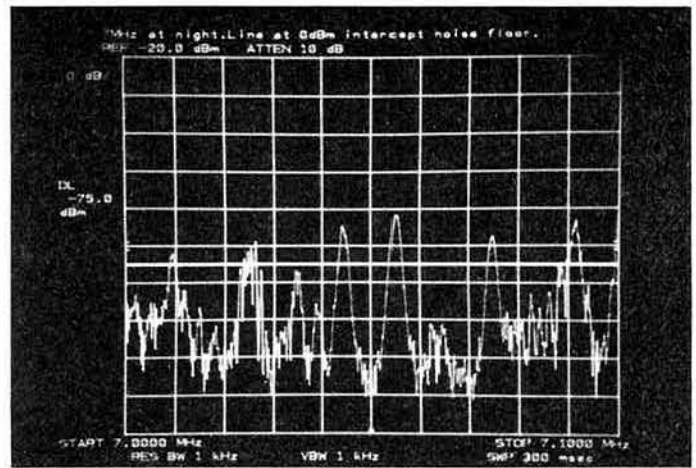


Fig 9. 7MHz at night. Line at noise level given by 0dBm intercept point

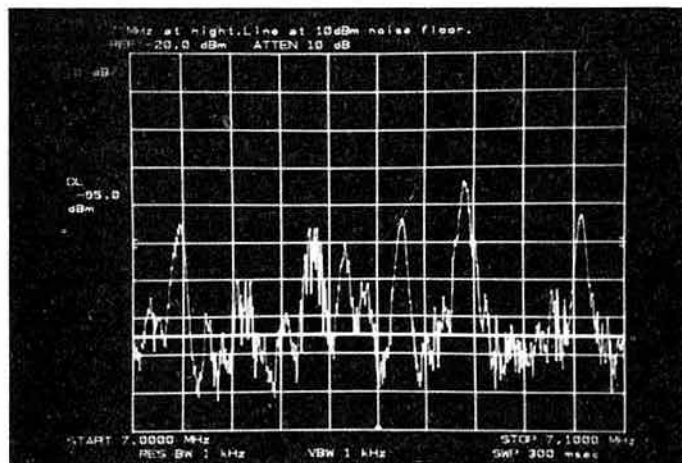


Fig 10. 7MHz at night. Line at noise level given by +10dBm intercept point

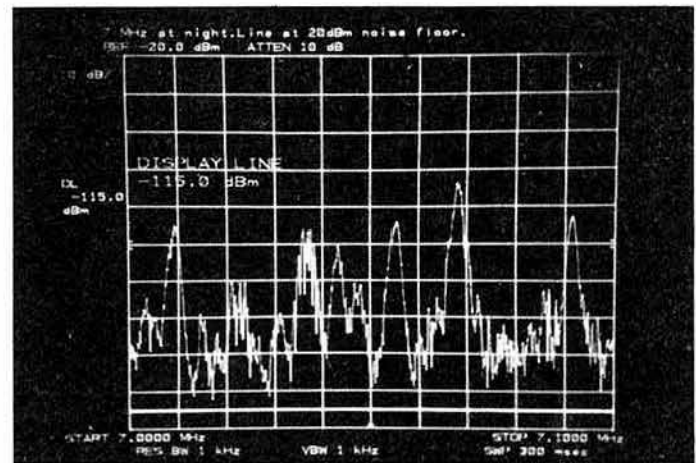


Fig 11. 7MHz at night. Line at noise level given by +20dBm intercept point

of the 7MHz band, which would get into the front-end. Improving the intercept point by 20dBm would reduce the worst-case imd by 40dB, and the resulting signals would probably be equal to the atmospheric noise level, provided that the reciprocal mixing performance was such that the 10 had noise sidebands at about -110dBc in a 1Hz bandwidth as close in as possible (see Figs 8-11). In [7], Barrs refers to a paper by Gardiner and Yousif entitled "Distortion effects in switching diode modulators arising from local oscillator interference", and the conclusion is drawn that there is in fact 6dB less noise than would be expected—at least in commutative diode mixers.

Techniques for reducing the effects of intermodulation have been examined in great detail in the literature, both professional and amateur, over the last few years, but only recently has the problem of imd in crystal filters been examined openly. Claims have been made for filters having +40dBm intercept points, but the method used for evaluation is considered very dubious by a number of authorities, and figures of +25 to +30dBm for carefully-chosen filters are more common. Average filters appear to run at about +13 to +20dBm for filters such as KVG XF9B types, and similarly for the majority of the 10·7MHz radiotelephone ones. Collins mechanical filters are much worse than this according to the data sheet, although this does not specifically state the levels for signals at 5 and 10, or 10 and 20kHz removed from the centre of the passband, which is where the measurements need to be made. It has also been claimed that the removal of ferrite-cored transformers from the filter input circuitry removes the imd problem, but this suggests not only bad design, insofar as the flux density is too high, but fails to explain why ladder filters without any transformers at all are just as bad! The most convincing explanation that the author has heard is that the effect is caused by the stressing of the crystal so that Hooke's Law of deformation no longer applies, but cleanliness of crystal manufacture has a lot to do with the results. Indeed, one prominent professional manufacturer who selects filters includes the direction of signal flow, as some filters are often much better one way round than the other.

For some time, mixers have been under scrutiny, but there are a number

of mixers of various types now available in a large number of varieties—diode, transistor, fet and ic—capable of giving intercept points of +20dBm or more. Here again it should be noted that gain and intercept points can be interdependent insofar as output intercepts can be quoted for mixers with gain, and input intercepts for those with loss, in order to play specmanship. In this respect: *caveat emptor*—"let the buyer beware!" Generally speaking, commutative diode mixers require an accurate wideband termination if they are to provide the best performance, and the calculations need to be done with great care if overall system performance is to be obtained. See [8, 9 and 10].

The advent of the synthesized receiver has led to the problems of reciprocal mixing becoming apparent, and this is likely to increase. Frequency synthesizers are popular for their ease of use, with the capabilities of programming, memories and so on, but pay the price in terms of reciprocal mixing. Single-loop synthesizers providing small steps by the use of a low reference frequency tend to be very poor in this respect, unless the tuning range of the vco is very restricted, because of the low loop bandwidth which is necessary. It is fair to say that single-loop phase-locked-loop synthesizers with low reference frequencies are best avoided in high-performance receivers—with the notable exception of the Racal Digiphase variety, which is not likely to be generally used in amateur equipment, and in any case has a high reference frequency.

In general the home constructor almost certainly produces a much better receiver by avoiding synthesizers completely at hf. The designs published in various magazines around the world have not yet shown phase noise performances which are good enough for an advanced amateur receiver, and considering the time and trouble involved in home-constructing a receiver, it may as well be spent in getting good performance as bad. In addition, the necessity for a synthesizer in terms of frequency accuracy, resettability, long-term stability or remote control does not exist in the vast majority of amateur stations in such a degree that two vfos will not suffice. Finally, even at today's prices, the cost of 12 quartz crystals plus a vfo and filters is likely to be cheaper than a synthesized injection system.

Summary

The importance of not only intermodulation but reciprocal mixing has been studied, with its relationship to the dynamic range of the receiver. Dynamic range has not been defined in quantitative terms except for intermodulation, and the requirements of modern amateur receivers in respect of intermodulation and reciprocal mixing have been evaluated, with reference to their likely rf environment. It is hoped that some idea of the importance of these parameters and their effects in terms of signals "not heard" has been given, and that more understanding of the parameters so carefully measured by reviewers will be gained.

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TECHNICAL TOPICS

(Continued from page 222)

shortage of young engineers with specialist training in analogue rf design and rf propagation. This results from students and teachers preferring the mathematical certainties of digital electronics to the more vague, but often more challenging, analogue systems—a tendency that has probably been encouraged by the higher initial salaries associated with the digital and computing side of the electronics industry.

A further problem is that fundamental rf propagation and research projects often involve time scales appreciably longer than can be fitted into normal courses. One sees very little evidence these days of much professional research in the UK into hf propagation, most of the work is on vhf, uhf and shf propagation. Yet, despite the increasing amount of digital technology now to be found in amateur radio stations (nineteenth-century telegraphy was essentially an early form of digital systems), the hobby is still largely dependent on analogue circuit design and the vagaries of hf and vhf propagation. Perhaps one of the contributions that the hobby can continue to make to the science and craft of radio communication is to keep alive interest in these areas.

Andrew Churchley, G4EAQ, described in *Radio Communication* February 1984 an interesting digital technique permitting the use of a home computer to work out the beam patterns from phased vertical antennas. It is perhaps worth recalling that almost 40 years ago the RCA antenna engineering team led by Dr George Brown developed an ingenious instrument they called an "antennalyzer" which, simply by setting 16 knobs with amplitude and phase information, displayed the radiation pattern of a complex vertical array on a large-screen oscilloscope. The "antennalyzer" needed a large enclosed rack several feet high, but nevertheless it formed an extremely flexible "analogue computer" that permitted rapid design of complex directional arrays with accurately-placed deep nulls as required in mf broadcast practice.

That analogue circuit design is in something of a decline at the moment is suggested by the disappearance from recent issues of the fortnightly *Electronics* magazine, one of the leading journals of the industry, of such features as "Designers' Casebook", the source of a considerable number of innovative circuits that have appeared over past decades in *TT*. Similarly *Electronic Design* has become increasingly orientated towards data processing and computer electronics.

Radio frequency design

More and more the publishing world reflects this concentration on data and digits. "Numerous books are available at all levels for the student of digital methods, especially those methods concerned with microprocessors. There is less information available for the beginner rf engineer other than sophisticated graduate-level treatments."

The above is a quotation from the preface to *Introduction to Radio Frequency Design*, a 1982 book which George May, G4RZF, feels should be drawn to the attention of *TT* readers. There is another excellent reason for following his advice: the author is Wes Hayward, W7ZOI, of Tektronix

who, apart from innumerable *QST* articles on simple hf receivers etc, is co-author (with Doug DeMaw, W1FB) of that bible for experimenters: *Solid State Basics*.

His latest book, it must be emphasized, is not intended primarily for the radio amateur, other than those with a moderate formal background in electronic engineering, but rather for students and those working professionally in the industry. Nevertheless, as he points out, "there are many serious experimenters with an interest in the subject matter". He stresses that "a central theme employed throughout the book is simplicity. I have observed that a thorough and complete analysis of a circuit using the simplest of circuit models is a pre-requisite to the level of understanding that leads to innovation."

G4RZF was much impressed by the book, and has reviewed it for his local club (Swindon and District 'Microcircuit') noting particularly four of the eight chapters: two-port networks; practical amplifiers and mixers; oscillators and frequency synthesizers; and the receiver as an rf system. I would add that the earlier chapters also include some valuable material on the design of crystal filters (including ladder filters), large signal operation of bipolar transistors, and a chapter on transmission lines including the use of the Smith Chart. His explanation of distortion in amplifiers includes the concept of intercept points; he also rightly stresses the importance of oscillator noise, often disregarded in the traditional textbooks.

So, this is *not* a book for those who enjoy and are content with building their own equipment from published designs, or who seek entirely non-mathematical explanations. But, since it escaped the net of the *Rad Com* "book review" column in 1982, I take this opportunity of endorsing G4RZF's view that it well worth the attention of the more "professional amateurs", even if it is a question of borrowing rather than buying a copy. (*Introduction to Radio Frequency Design*, W. H. Hayward. Published in the UK by Prentice/Hall International, first edition 1982, 384 + xiv pages, ISBN: 13-494021-0, £20.95).

Tips and topics

A novel form of probe for identifying faulty, in-operative components is being marketed in the USA by Metrifact. It operates by sensing very small temperature changes when held above semiconductor components, resistors etc. The "Thermoprobe" is claimed to be able to detect a temperature change of only one-fortieth of a degree Centigrade when the probe is held about one-sixteenth of an inch above a component. The meter on the handunit is first set to a null position representing ambient temperature. It uses a thermistor probe in a modified Wheatstone-bridge circuit.

Mike White, G4IPY, had a go at G3UUR's point about reflected "power" in transmission lines (October), but I believe he has missed the point that G3UUR was making. By definition, power represents energy actually consumed in a load; until then it is strictly speaking *energy*, but it is *real* energy sure enough! In effect, G3UUR was quibbling over the semantics and not with W2DU's basic premise. Very few of us bother to distinguish between power and energy as it does not affect matters in a practical sense. □